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Applicants:	Masahiro Ishida, et al.
Serial No.:	09/882,290
Filing Date:	June 14, 2001
Title:	Apparatus for and Method of Measuring Jitter

October 10, 2001
San Francisco, California

Commissioner of Patents
Washington, DC 20231

PRELIMINARY AMENDMENT

Sir:

Enclosed herewith are four sets of pages: (1) pages with revision marks showing amendments to the specification, (2) pages containing substitute paragraphs in clean form, (3) a revised drawing showing changes to Figure 6 in red, and (4) a substitute drawing for Fig. 6.

Preliminary to the examination of this application, please amend the application as shown below.

AMENDMENTS

In the Specification

Please amend the specification as shown by revision marks on the enclosed set of pages. A separate set of pages contains the amended paragraphs in clean form. Amendments to the specification are found at the following locations:

On page 4, line 18, replace "zero-crossing method" with --time interval analyzer method--.

On page 5, lines 2-3, replace "as adaptive zero-crossing points approximation" with -- as approximated zero-crossing points). --.

On page 22, lines 26-27, after "value of" delete "a real part x(t) of an analytic signal of".

On page 23, line 15, replace "a real part x(t) of an analytic signal." with --an inputted signal under measurement. --.

On page 31, line 21, replace "δ [k]" with --J[k] --.

On page 31, line 24, replace "106" with -- 105 --.

On page 32, line 20, after " value of", insert -- cycle-to-cycle --.

On page 32, line 21, after " value of", insert -- cycle-to-cycle --.

On page 32, line 23, replace "instantaneous period sequence." with --cycle-to-cycle period jitter sequence --.

On page 38, line 27, replace "controller 1302" with -- limiter 1302 --.

On page 39, line 4, replace "controller 1302" with -- limiter 1302 --.

In the Drawings

A proposed amendment to Fig. 6 is shown in red on the enclosed drawing. A drawing incorporating this amendment is also enclosed.

DISCUSSION

The amendments to the specification are being made to correct errors in the English language translation.

The amendment to Fig. 6 is being made in order to conform the drawing to the text of the specification.

Respectfully submitted,

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Certificate of Mailing Under 37 CFR 1.8

I certify that this Preliminary Amendment and any enclosed materials are being deposited with the United States Postal Service on October 10, 2001 with sufficient postage as first class mail in an envelope addressed to Commissioner for Patents, Washington, DC 20231.

David N. Lathrop

Enc. Set of pages with revision marks showing amendments to the specification
Set of pages containing the amended paragraphs in clean form
Revised drawing showing changes to Figure 6 in red
Substitute drawing for Fig. 6
Postcard

1. PAGES SHOWING REVISIONS TO SPECIFICATION

$$\Delta\phi_{RMS}=\sqrt{\frac{1}{N}\sum_{k=0}^{N-1}\Delta\phi^2[n]}$$

This method is referred to as the Δϕ method, since a peak value of timing jitter (peak-to-peak value) and a root-mean-square value of timing jitter are obtained as described above from the instantaneous phase noise waveform Δϕ(t).

According to the Δϕ method, a timing jitter can be measured at high speed with relatively high accuracy.

Since the jitter measurement method in the time interval analyzer system includes an intermediate dead-time during which no measurement can be performed after each one of periodic measurements is performed, there is a problem that it takes a long time to acquire a number of data that are required for a histogram analysis.

In addition, in a jitter measurement method in which a wide-band oscilloscope and an interpolation method are combined, there is a problem that a jitter is overestimated (overestimation). That is, the measured jitter values in this method are not compatible with the values measured by the ~~zero-crossing method~~time interval analyzer method. For example, a jitter measurement result by the time interval analyzer and a jitter measurement result by the interpolation method for a 400 MHz clock signal are shown in Figs 7A and 7B, respectively so that those measurement results can be compared with each other. According to those figures, the measured value by the time interval analyzer is 7.72 ps (RMS), while the measured value by the interpolation method is 8.47 ps (RMS), and this value is overestimated. In addition, the jitter measurement method by the interpolation method requires a long measurement time.

signal under measurement is approximated by a sampling point closest to each zero-crossing point (this point is referred to ~~as adaptive zero-crossing points approximation~~as approximated zero-crossing points). Therefore, there is a possibility that an amount of error in jitter measurement is increased when an over-sampling rate in the signal digitization process is small.

It is an object of the present invention to provide a jitter measurement apparatus and its method that can estimate a jitter value compatible with a jitter value measured by the conventional time interval analyzer method.

SUMMARY OF THE INVENTION

The jitter measurement apparatus according to the present invention comprises: phase error estimation means for estimating sampling points close to corresponding zero-crossing points of the signal under measurement as approximated zero-crossing points, for obtaining phase errors between the approximated zero-crossing points and the corresponding zero-crossing points of the signal under measurement, and for outputting a zero crossing phase error data sequence; and a period jitter estimation means for obtaining a period jitter sequence of the signal under measurement from the phase error data sequence.

It is desirable that the jitter measurement apparatus according to the present invention further comprises: cycle-to-cycle period jitter estimation means to which the period jitter sequence is inputted for calculating its difference sequence, and for outputting a cycle-to-cycle period jitter sequence.

It is desirable that the jitter measurement apparatus according to the present invention further comprises: band-pass filter means for passing therethrough predetermined frequency components of the signal under

oscilloscope is used for the digitization of the analog signal.

In addition, in the jitter measurement method according to the present invention, a period jitter can also be measured with high accuracy by removing amplitude modulation (AM) components of the signal under measurement using waveform clipping means in the state that phase modulation (PM) components corresponding to jitter are retained.

BAND-PASS FILTER

A band-limitation of a digitized digital signal can be realized by a digital filter, and can also be performed by Fourier transform. Next, a band-pass filtering using FFT (Fast Fourier Transform) will be explained below. FFT is a method of transforming at high speed a signal waveform in time domain to a signal in frequency domain.

First, a digitized signal under measurement $x(t)$ shown in Fig. 16 is transformed to a signal $X(f)$ in frequency domain using FFT. Fig. 17 shows a power spectrum of the transformed signal $X(f)$. Then data around a fundamental frequency (400 MHz in this example) of the signal $X(f)$ are retained and the other data are made zero. Fig. 18 shows a signal $X_{BP}(f)$ of the band-limited signal in frequency domain. Finally, a band-limited signal waveform $x_{BP}(t)$ in time domain can be obtained by applying inverse FFT to the band-limited signal $X_{BP}(f)$ in frequency domain. Fig. 19 shows the band-limited signal waveform $x_{BP}(t)$ in time domain.

APPROXIMATED ZERO-CROSSING POINT DETECTION METHOD

Next, an approximated zero-crossing point detection method will be described. First, the maximum value of ~~a real part x(t) of an analytic signal of~~ an inputted signal under measurement is defined as 100% level, and the

minimum value is defined as 0% level to calculate 50% level signal value V (50%) as a zero crossing level. Then, a difference between a sample value x(j-1) and the 50% level V (50%) of the x(t) and a difference between its adjacent sample value x(j) and the 50 % level V(50%) of the x(t), i.e., (x(j-1)-V(50%)) and (x(j)-V(50%)) are calculated, and furthermore a product of those difference values (x(j-1)-V(50%))× (x(j)-V(50%)) is calculated. When the x(t) crosses 50% level, i.e., zero-crossing level, the sign of the difference between the sample value and the V(50%), i.e., (x(j-1) –V(50%)) or (x(j)-V(50%)) changes from a negative sign to a positive sign or from a positive sign to a negative sign. Therefore, when the product is negative, it is detected that the x(t) has passed the zero-crossing level, and a time point j-1 or j, at which a smaller absolute value of the difference between the sample value and the V (50%), i.e., (x(j-1) –V(50%)) or (x(j)-V(50%)) is detected, is obtained as an approximated zero crossing point. Fig. 20 shows a waveform of ~~a real part x(t) of an analytic signal~~an inputted signal under measurement~~.~~ The marks ○ in Fig. 20 indicate the detected points (approximated zero-crossing points) closest to the corresponding rising zero-crossing points.

WAVEFORM CLIPPING

Waveform clipping means removes AM components from an input signal and retains only PM components in the input signal. A waveform clipping is performed by: 1) multiplying an analog or digital input signal by a constant, 2) replacing a signal value greater than a predetermined threshold value Th1 with the threshold value Th1, and 3) replacing a signal value less than a predetermined threshold value Th2 with the threshold value Th2. Here, it is assumed that the threshold value Th1 is greater than the threshold value Th2. Fig. 21 shows a clock signal having AM components. Since an

period jitter of the signal under measurement, the peak-to-peak detector 107 obtains the peak-to-peak value of period jitter using the equation (29), the RMS detector 108 obtains the RMS value of period jitter using the equation (28), and the histogram estimator 109 obtains the histogram from the instantaneous period waveform data. In order to estimate the approximated zero-crossing points, as apparent from the explanation of "approximated zero-crossing point detection method" in the aforementioned section of summary of the invention, the signal to which a band-pass filtering process has been applied must be an over-sampled (sampling at a frequency equal to or higher than three times of the fundamental frequency) digital signal. If the output of the band-pass filter 103, i.e., the band-limited signal in the step 202 is an analog signal, the signal inputted to the phase error estimator 104 is converted into a digital signal within the phase error estimator 104, whereby the digital signal is processed. That is, in the step 203, the signal is processed after the input signal is converted into a digital signal.

Fig. 25 shows another example of configuration of a jitter measurement apparatus used in the embodiment of the present invention. This jitter measurement apparatus 300 is same as the jitter measurement apparatus shown in Fig. 23 except that the jitter measurement apparatus 300 includes a cycle-to-cycle period jitter estimator 301 to which the period jitter sequence ~~δ[k]~~J[k] is inputted for calculating a difference waveform using the equation (30) to output a cycle-to-cycle period jitter sequence $J_{CC}[k]$, and a switch 302 for selecting an output sequence from the period jitter estimator ~~106~~105 or the cycle-to-cycle period jitter estimator 301. For simplicity, the explanation of the overlapped portion is omitted. In this case, the jitter detector 106 obtains a jitter value (a peak-to-peak value, an RMS value or a histogram) of the signal under measurement from the period jitter sequence

J[k] or the cycle-to-cycle period jitter sequence $J_{CC}[k]$.

Next, the operation in the case where a jitter measurement of the signal under measurement is performed using the jitter measurement apparatus 300 according to the present invention will be explained below. Fig. 26 shows another processing procedure of a jitter measurement method according to the present invention. This jitter measurement method is same as the jitter measurement method shown in Fig. 24 except that this jitter measurement method includes: a step 401 for calculating, from the period jitter sequence J[k] estimated by the period jitter estimator 105, a difference waveform of the period jitter sequence J[k] by means of the cycle-to-cycle period jitter estimator 301, after the period jitter of the signal under measurement is obtained from the period jitter sequence J[k], and for obtaining a cycle-to-cycle period jitter sequence $J_{CC}[k]$; and a step 402 for obtaining a cycle-to-cycle period jitter of the signal under measurement from the cycle-to-cycle period jitter sequence $J_{CC}[k]$ in the state that the jitter detector 106 connects the switch 302 to the cycle-to-cycle period jitter estimator 301 side. In this case, the explanation of the overlapped portion is omitted. In the step 402 for obtaining the cycle-to-cycle period jitter of the signal under measurement, the peak-to-peak detector 107 obtains a peak-to-peak value of cycle-to-cycle period jitter using the equation (32), the RMS detector 108 obtains an RMS value of cycle-to-cycle period jitter using the equation (31), and the histogram estimator 109 obtains a histogram from the ~~instantaneous period sequence~~cycle-to-cycle period jitter sequence.

The jitter measurement apparatus shown in Fig. 25 can also be constructed as an apparatus for estimating only cycle-to-cycle period jitter. In this case, the switch 302 for selecting a sequence can be omitted. Similarly, the jitter measurement method shown in Fig. 26 may also estimate

calculation of the equation (9) from the instantaneous angular frequency sequence ω[k] and the fundamental period T_0.

Next, the operation in the case of performing a period jitter estimation of the signal under measurement using this period jitter estimator 105 shown in Fig. 33 will be explained. Fig. 34 shows a processing procedure of this period jitter estimation method. First, in step 1201, the instantaneous angular frequency estimator 1101 obtains the instantaneous angular frequency sequence ω[k] of the signal under measurement from the zero-crossing phase error data δ[k] estimated by the zero-crossing phase error estimator 104 and the zero-crossing time interval sequence $T_{k,k+1}$ obtained from the zero-crossing sampler 501. Next, in step 1202, the period jitter calculator 1102 obtains the period jitter sequence J[k] of the signal under measurement from the instantaneous angular frequency sequence ω[k] obtained from the instantaneous angular frequency estimator 1101 and the fundamental period T_0 stored in the memory 102, and the process ends. In the step 1201 for obtaining the instantaneous angular frequency sequence ω[k] of the signal under measurement, the instantaneous angular frequency estimator 1101 obtains the instantaneous angular frequency using the equation (10). In addition, in the step 1202 for obtaining the period jitter J[k] of the signal under measurement from the instantaneous angular frequency sequence ω[k], the period jitter calculator 1102 obtains the period jitter sequence J[k] using the equation (9).

Fig. 35 shows an example of configuration of the band-pass filter 103 used in the jitter measurement apparatus 100. This band-pass filter 103 comprises: a time domain to frequency domain transformer 1301 for transforming the signal under measurement into a signal in frequency domain; a bandwidth ~~controller~~limiter 1302 for extracting only components around the

fundamental frequency of the signal under measurement from the output of the time domain to frequency domain transformer 1301; and a frequency domain to time domain transformer 1303 for inverse-transforming the output of the bandwidth ~~controller~~limiter 1302 into a signal in time domain. The time domain to frequency domain transformer 1301 and the frequency domain to time domain transformer 1303 may be packaged using FFT and inverse FFT, respectively.

Next, the operation in the case of performing a band-limitation of the signal under measurement using this band-pass filter 103 shown in Fig. 35 will be explained. Fig. 36 shows a processing procedure of this band-limitation method. First, in step 1401, the time domain to frequency domain transformer 1301 applies FFT to the signal under measurement to transform a signal in time domain into a signal in frequency domain. Next, in step 1402, the bandwidth limiter 1302 retains the components around the fundamental frequency of the signal under measurement in the transformed signal in frequency domain, and replaces the other frequency components with zero to band-limit the signal in frequency domain. Finally, in step 1403, the frequency domain to time domain transformer 1303 applies inverse FFT to the band-limited signal in frequency domain to transform the signal in frequency domain into a signal in time domain, and the process ends.

Fig. 37 shows another example of configuration of the band-pass filter 103 used in the jitter measurement apparatus 100. This band-pass filter 103 comprises: a buffer memory 1501 for storing therein the signal under measurement; a data selector 1502 for extracting the signal in the sequential order from the buffer memory 1501 such that the signal being extracted is partially overlapped with the signal extracted just before; a window function multiplier 1503 for multiplying each extracted partial signal by a window function; a time domain to frequency domain transformer 1504 for

2. PAGES CONTAINING SUBSTITUTE PARAGRAPHS IN CLEAN FORM

Substitute text for paragraph at page 4, line 14:

In addition, in a jitter measurement method in which a wide-band oscilloscope and an interpolation method are combined, there is a problem that a jitter is overestimated (overestimation). That is, the measured jitter values in this method are not compatible with the values measured by the time interval analyzer method. For example, a jitter measurement result by the time interval analyzer and a jitter measurement result by the interpolation method for a 400 MHz clock signal are shown in Figs 7A and 7B, respectively so that those measurement results can be compared with each other. According to those figures, the measured value by the time interval analyzer is 7.72 ps (RMS), while the measured value by the interpolation method is 8.47 ps (RMS), and this value is overestimated. In addition, the jitter measurement method by the interpolation method requires a long measurement time.

Substitute text for paragraph at page 4, line 26:

In addition, in the conventional $\Delta\phi$ method, a zero-crossing point of a signal under measurement is approximated by a sampling point closest to each zero-crossing point (this point is referred to as approximated zero-crossing points). Therefore, there is a possibility that an amount of error in jitter measurement is increased when an over-sampling rate in the signal digitization process is small.

Substitute text for paragraph at page 22, line 25:

Next, an approximated zero-crossing point detection method will be described. First, the maximum value of an inputted signal under measurement is defined as 100% level, and the minimum value is defined as 0% level to calculate 50% level signal value V (50%) as a zero crossing level. Then, a difference between a sample value x(j-1) and the 50% level V (50%) of the x(t) and a difference between its adjacent sample value x(j) and the 50 % level V(50%) of the x(t), i.e., (x(j-1)-V(50%)) and (x(j)-V(50%)) are calculated, and furthermore a product of those difference values (x(j-1)-V(50%)) × (x(j)-V(50%)) is calculated. When the x(t) crosses 50% level, i.e., zero-crossing level, the sign of the difference between the sample value and the V(50%), i.e., (x(j-1) –V(50%)) or (x(j)-V(50%)) changes from a negative sign to a positive sign or from a positive sign to a negative sign. Therefore, when the product is negative, it is detected that the x(t) has passed the zero-crossing level, and a time point j-1 or j, at which a smaller absolute value of the difference between the sample value and the V (50%), i.e., (x(j-1) –V(50%)) or (x(j)-V(50%)) is detected, is obtained as an approximated zero crossing point. Fig. 20 shows a waveform of an inputted signal under measurement. The marks ○ in Fig. 20 indicate the detected points (approximated zero-crossing points) closest to the corresponding rising zero-crossing points.

Substitute text for paragraph at page 31, line 16:

Fig. 25 shows another example of configuration of a jitter measurement apparatus used in the embodiment of the present invention. This jitter measurement apparatus 300 is same as the jitter measurement apparatus shown in Fig. 23 except that the jitter measurement apparatus 300 includes a cycle-to-cycle period jitter estimator 301 to which the period jitter sequence J[k] is inputted for calculating a difference waveform using the equation (30) to output a cycle-to-cycle period jitter sequence $J_{CC}[k]$, and a switch 302 for selecting an output sequence from the period jitter estimator 105 or the cycle-to-cycle period jitter estimator 301. For simplicity, the explanation of the overlapped portion is omitted. In this case, the jitter detector 106 obtains a jitter value (a peak-to-peak value, an RMS value or a histogram) of the signal under measurement from the period jitter sequence J[k] or the cycle-to-cycle period jitter sequence $J_{CC}[k]$.

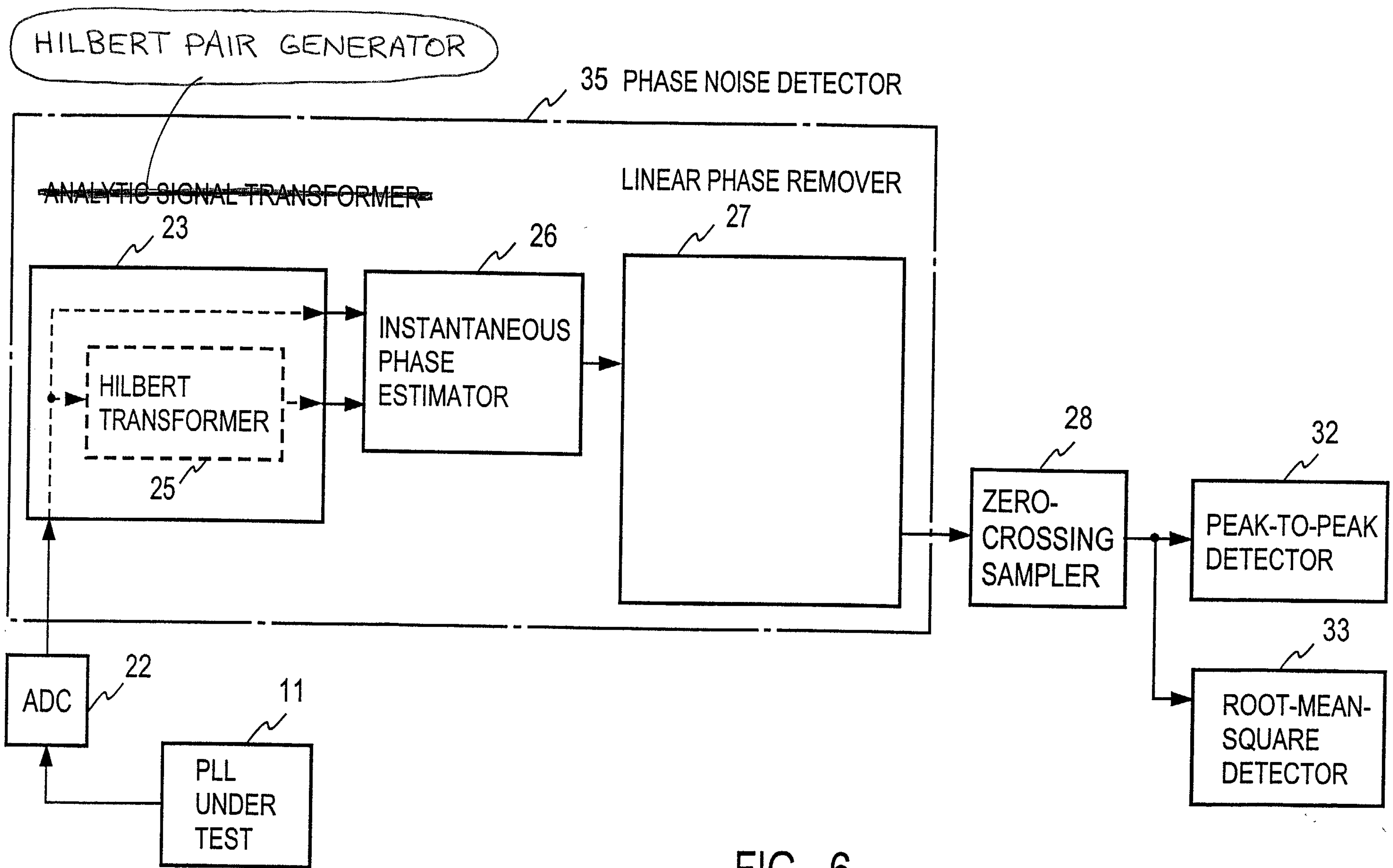
Substitute text for paragraph at page 32, line 2:

Next, the operation in the case where a jitter measurement of the signal under measurement is performed using the jitter measurement apparatus 300 according to the present invention will be explained below. Fig. 26 shows another processing procedure of a jitter measurement method according to the present invention. This jitter measurement method is same as the jitter measurement method shown in Fig. 24 except that this jitter measurement method includes: a step 401 for calculating, from the period jitter sequence J[k] estimated by the period jitter estimator 105, a difference waveform of the period jitter sequence J[k] by means of the cycle-to-cycle period jitter estimator 301, after the period jitter of the signal under measurement is obtained from the period jitter sequence J[k], and for obtaining a cycle-to-cycle period jitter sequence $J_{CC}[k]$; and a step 402 for obtaining a cycle-to-cycle period jitter of the signal under measurement from the cycle-to-cycle period jitter sequence $J_{CC}[k]$ in the state that the jitter detector 106 connects the switch 302 to the cycle-to-cycle period jitter estimator 301 side. In this case, the explanation of the overlapped portion is omitted. In the step 402 for obtaining the cycle-to-cycle period jitter of the signal under measurement, the peak-to-peak detector 107 obtains a peak-to-peak value of cycle-to-cycle period jitter using the equation (32), the RMS detector 108 obtains an RMS value of cycle-to-cycle period jitter using the equation (31), and the histogram estimator 109 obtains a histogram from the cycle-to-cycle period jitter sequence.

Substitute text for paragraph at page 38, line 23:

Fig. 35 shows an example of configuration of the band-pass filter 103 used in the jitter measurement apparatus 100. This band-pass filter 103 comprises: a time domain to frequency domain transformer 1301 for transforming the signal under measurement into a signal in frequency domain; a bandwidth limiter 1302 for extracting only components around the fundamental frequency of the signal under measurement from the output of the time domain to frequency domain transformer 1301; and a frequency domain to time domain transformer 1303 for inverse-transforming the output of the bandwidth limiter 1302 into a signal in time domain. The time domain to frequency domain transformer 1301 and the frequency domain to time domain transformer 1303 may be packaged using FFT and inverse FFT, respectively.

3. REVISED DRAWING SHOWING CHANGES TO FIGURE 6



HILBERT PAIR GENERATOR
35 PHASE NOISE DETECTOR
~~ANALYTIC SIGNAL TRANSFORMER~~
23
HILBERT TRANSFORMER
25
26
INSTANTANEOUS PHASE ESTIMATOR
LINEAR PHASE REMOVER
27
28
ZERO-CROSSING SAMPLER
32
PEAK-TO-PEAK DETECTOR
33
ROOT-MEAN-SQUARE DETECTOR
ADC
22
11
PLL UNDER TEST

FIG. 6

4. SUBSTITUTE DRAWING FOR FIGURE 6

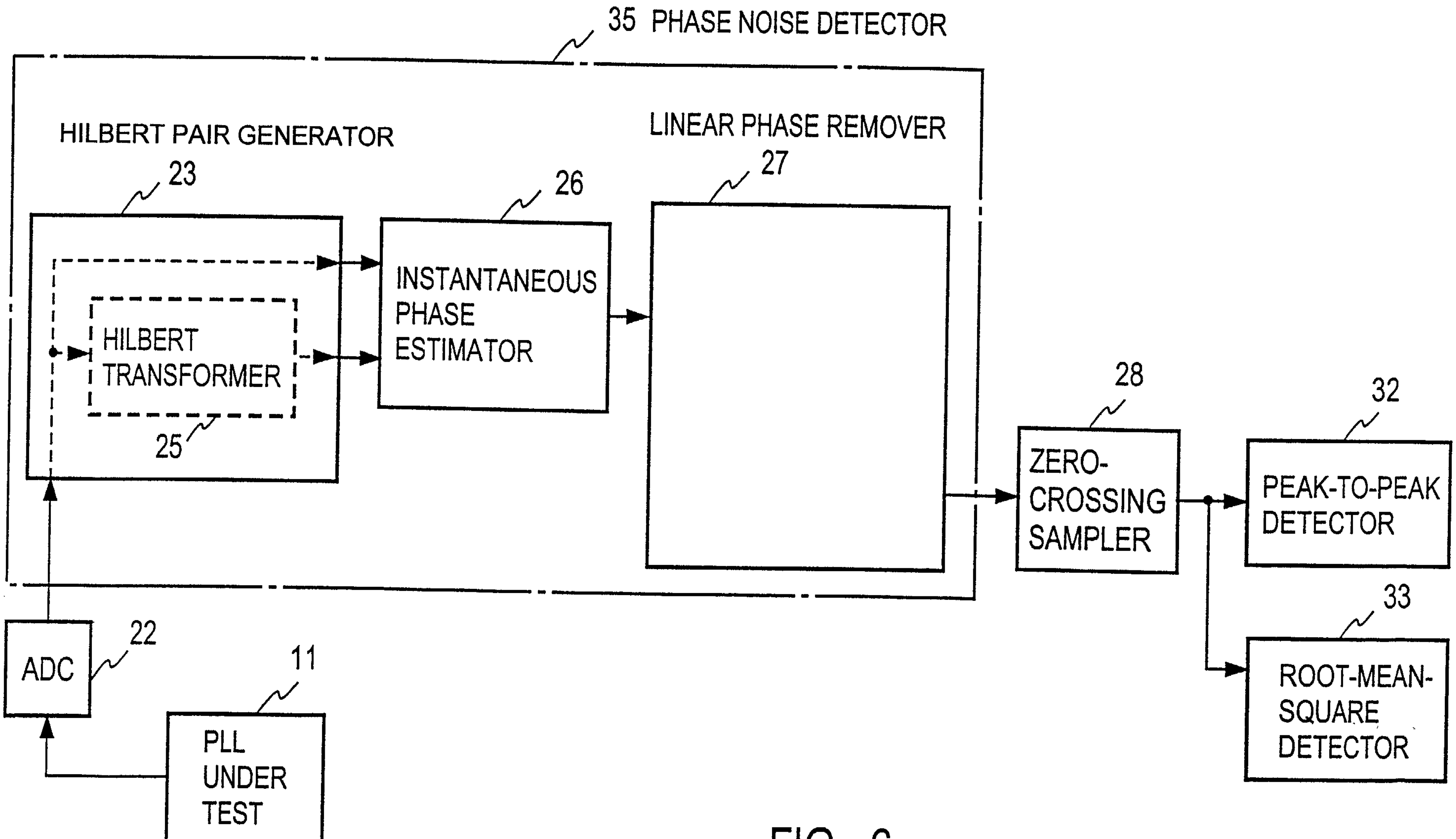

35 PHASE NOISE DETECTOR
HILBERT PAIR GENERATOR
23
HILBERT TRANSFORMER
25
26
INSTANTANEOUS PHASE ESTIMATOR
LINEAR PHASE REMOVER
27
28
ZERO-CROSSING SAMPLER
32
PEAK-TO-PEAK DETECTOR
33
ROOT-MEAN-SQUARE DETECTOR
ADC
22
11
PLL UNDER TEST

FIG. 6